# A MODIFIED SERRODYNE TECHNIQUE TO PRODUCE PHASE MODULATION HAVING EQUIVALENT HIGH-MODULATION INDEX

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# THESIS

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A Modified Serrodyne Technique to

Produce Phase Modulation Having Equivalent

High-Modulation Index

by

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# ABSTRACT

If a sawtooth wave is used as the modulating signal for the accelerating voltage of a traveling-wave tube, the input radio-frequency signal is phase modulated to produce a frequency shift in the output signal. The use of a sine wave as the modulating signal results in the familiar sine-wave-phase-modulated carrier as the output; however, it is difficult to obtain a large phase-modulation index without objectionable amplitude modulation. Specifically, if the peak-to-peak voltage swing of the modulating wave produces a phase shift of more than 2π radians in the radio-frequency carrier, the amount of amplitude modulation becomes severe. To overcome this severe amplitude modulation and yet produce large phase deviations in the carrier signal, i.e., broadband phase modulation, it is possible to chop the modulating waveform just as a ramp is chopped into a sawtooth. This thesis presents a design for so modifying the modulating signal be it sine wave, sawtooth, exponential or complex, and shows the resultant output spectrum of a traveling-wave tube whose carrier has been modulated by a chopped sine wave.

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# I. INTRODUCTION

Transit-time devices such as klystrons and traveling-wave tubes (TWT) have been used for several years in producing frequency translation of a radio-frequency carrier by varying the transit time of the electron beam within the device. Approximately linear variation of electron-beam transit time in a TWT has been reported by Dr. Raymond C. Cumming [1] through periodic modification of the electron beam accelerating potential. That periodic modification was achieved by applying a sawtooth wave to the helix, cathode or electron gun of the TWT.

Serrodyne modulation as described by Cumming employs a TWT, a radio-frequency generator and a sawtooth generator. Basically its operation is as follows: the TWT is a transit-time device, i.e., output cycles lag the corresponding input cycles by a given time delay. Physically this time delay is the result of the finite time it takes electron bunches to propagate the length of the TWT. If no disturbing influence affects the velocity of the electron bunches, then uniformly spaced electron bunches at the input will arrive uniformly spaced at the output. However, by disturbing the velocity of the electron bunches, the output frequency is altered; in particular if the alteration consists of linearly increasing or decreasing the transit time the output will be uniform but lower or higher in frequency respectively than the input. The means of uniformly lowering or raising the output frequency is achieved through the application of a sawtooth to the electron-beam voltage of the TWT.

Although any sawtooth voltage of sufficient magnitude will alter the output frequency, there are factors which make particular sawtooth characteristics desirable.

First, the TWT operates most efficiently over a given beam-voltage range, i.e., the gain of the stage is a function of beam voltage. Hence it is obviously desirable to operate in the range which gives the greatest and most constant gain versus beam voltage to achieve greatest output power with least incidental amplitude modulation.

Second, during the flyback time of the sawtooth the output frequency is translated in opposite sense to the desired translation which occurs during the ramp portion of the sawtooth. Ideally then the flyback time should be zero.

Third, phase coherence should be maintained from one ramp period to the next ramp period of the sawtooth to obtain the largest output power at a single frequency. Thus it is desirable that the peak-to-peak sawtooth voltage cause an integral multiple of $2\pi$ radians phase shift in the output frequency.

Fourth, a controlled nonlinearity of the sawtooth ramp would be desirable to compensate for the square-root variation of electron velocity with beam voltage that exists in traveling wave tubes. Fortunately, since the fractional change in beam voltage is small, this controlled nonlinearity is not essential in most applications.

Thus an ideal sawtooth for serrodyne operation is one whose flyback time is zero and whose peak-to-peak voltage swing is that value which will cause precisely $2\pi$ radians phase shift in the output frequency. The ramp portion will be nonlinear to effect a linear change in electron beam velocity. Additionally, the grid voltage is varied simultaneously with the sawtooth ramp to maintain constant gain to eliminate amplitude modulation.

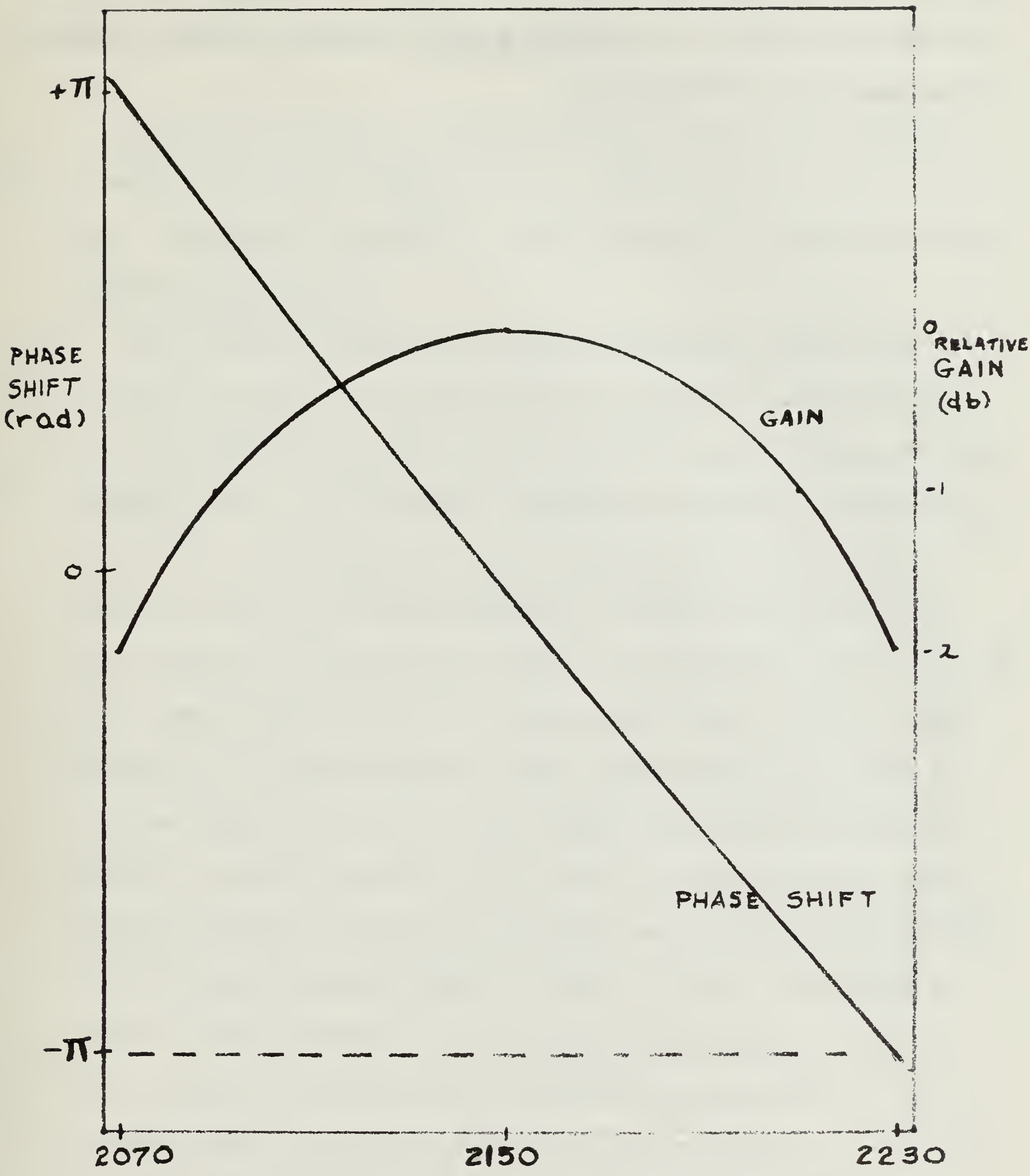
If such a sawtooth as just described were used to modulate the beam voltage of a TWT, the output frequency would be translated by the sawtooth frequency up or down from the input frequency depending upon whether the sawtooth caused an increase or decrease in electron beam velocity, respectively. If the sawtooth produced an output phase shift of $2\pi N$ where N is an integer then the frequency translation would be N times the sawtooth frequency, but with greater amplitude modulation than when N is unity.

Similarly for a modulating signal other than a sawtooth, the instantaneous output frequency shift is equal to the derivative of the instantaneous phase shift, and hence, approximately proportional to the derivative of the modulating waveform for beam voltage modulation. Specifically, if the modulating signal is sinusodial with time, then the instantaneous output frequency shift is proportional to the negative cosine.

Phase modulation has the property that the instantaneous phase of the modulated signal depends linearly upon the amplitude of the modulating signal, which is precisely what has been described above.

If the modulating signal were a complex wave such as an audio signal, the output of the TWT would be a carrier phase-modulated by the audio signal, and it could be detected to recover the useful intelligence. Unfortunately, the frequency deviation of the output would not be great enough to be of much practical value if the audio signal amplitude were only large enough to produce a $2\pi$-radian phase shift in the output signal. Moreover, if an attempt is made to increase the peak-to-peak excursions of the audio signal in order to produce larger phase shifts in the carrier, Fig. 1 clearly shows that undesirable amplitude modulation of the carrier will result. If the TWT were adjusted to give the flattest gain-versus-helix-voltage characteristic possible, and the modulating signal amplitude were increased to the point where an unacceptable amount of amplitude modulation of the carrier resulted, the frequency deviation of the carrier would still be rather small. Attendant with narrowband phase modulation are undesirable noise and interference characteristics, making it desirable to devise a means to process the audio signal in such a way that broader-band phase modulation could be achieved. Professor D. B. Hoisington conceived the means by combining the ideas of serrodyne modulation as explained earlier with the possibilities that are shown in the almost linear relationship of carrier phase shift versus helix voltage in Fig. 1. Basically what was needed was a means of chopping the modulating signal into segments just as the infinite nonlinear ramp (which would produce perfect serrodyne) was chopped into a sawtooth to produce an imperfect but useable carrier translation.

Thus the problem was quite similar to the one of producing an ideal sawtooth in that it was desirable to have a zero flyback time and a peak-to-peak voltage excursion giving an output phase shift of exactly $2\pi$ radians as explained earlier. However, the problem was also quite different in that with a sawtooth generator the output is repetitive and the flyback is always in one direction. The problem of chopping a complex waveform,on the other hand, requires a variable period and




PHASE SHIFT & GAIN vs. HELIX to CATHODE POTENTIAL
FIG. 1

both positive and negative flyback directions. The following sections describe the results of the author's work in devising suitable circuitry to accomplish the defined task.

## II. SYSTEM CONCEPT

As indicated earlier, what was needed was a means of chopping an audio signal into segments much as a ramp is chopped into a sawtooth. For ease of explanation a sine wave will be used as the audio signal in this section.

For a sine-wave input of arbitrary amplitude as shown in Fig. 2a the desired output for TWT beam modulation would be as depicted in Fig. 2b. Observe that between the flyback voltages $+kV_{\pi}$ and $-kV_{\pi}$ the waveform in Fig. 2b has the same instantaneous slope as the waveform in Fig. 2a. After amplification the chopped waveform would be as depicted in Fig. 2c. In these figures k is some positive fraction whose reciprocal represents the amount the chopped sine wave in Fig. 2b must be amplified to produce the chopped sine wave in Fig. 2c. The voltage $2V_{\pi}$ is the peak-to-peak voltage swing which will produce a $2\pi$ radian phase shift in the output signal of the TWT as discussed earlier. The most important point to note in these figures is that the time-rate-of-voltage change (slope)has been increased considerably in Fig. 2c over that which exists in Figs. 2a & 2b. If a signal as depicted in Fig. 2c were used as the modulating signal of a TWT the output signal of the TWT would be perturbed much more than it would be were a signal used of the form of that shown in Fig. 2a. The goal was earlier defined as modifying the modulating signal in a manner that would produce larger frequency deviations of the radio-frequency carrier than would be available if the modulating signal were not so modified. That the signal depicted in Fig. 2c would accomplish this goal should be clear. It should be equally clear that if a sine wave

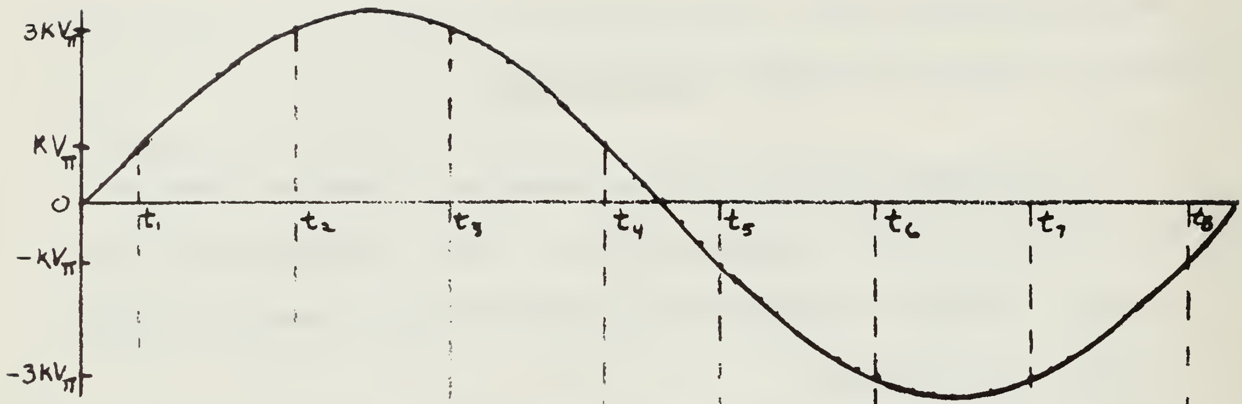



FIG. 2a SINE WAVE INPUT

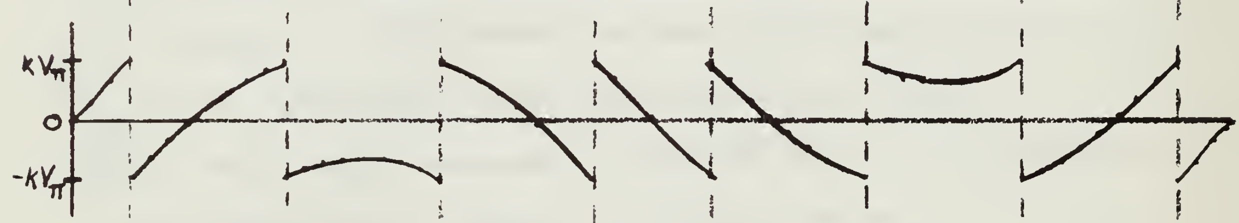



FIG. 2b. CHOPPED SINE WAVE

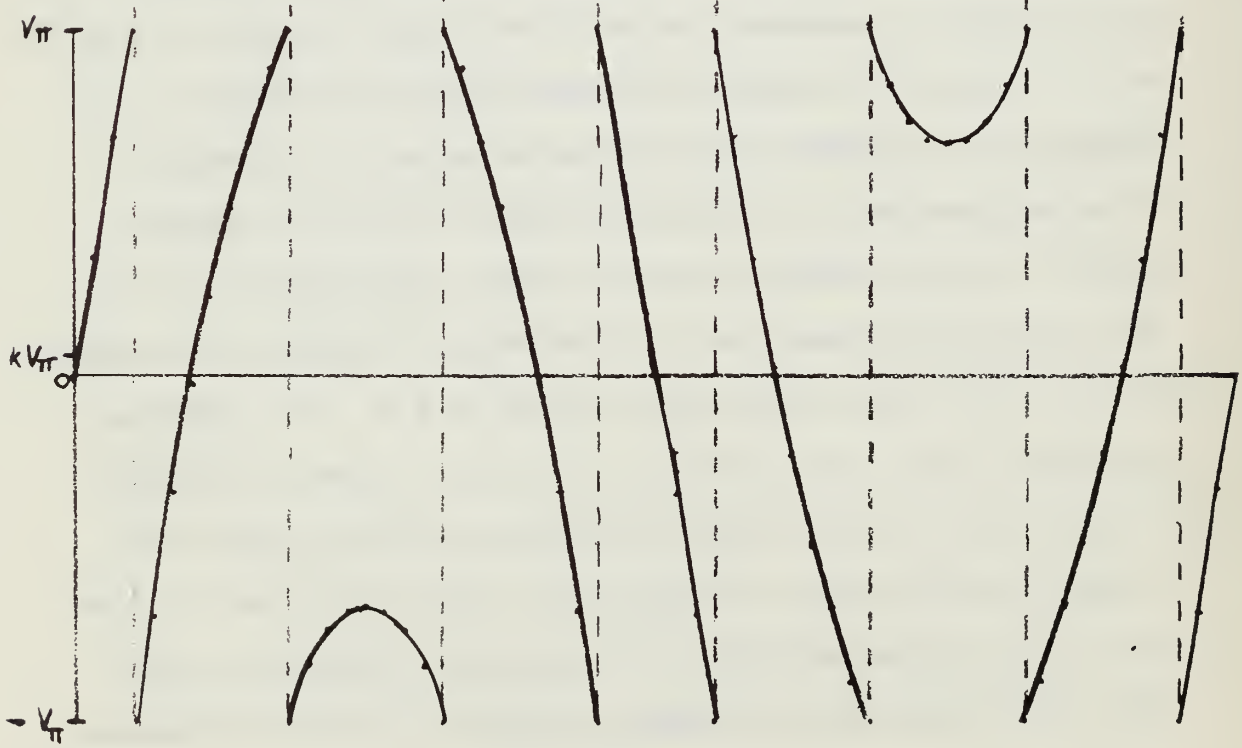



FIG. 2c AMPLIFIED CHOPPED SINE WAVE

were chopped into more segments and used as the modulating signal, the output spectrum of the TWT would contain even greater frequency deviations; i.e., the phase modulation index would be increased.

The block diagram shown in Fig. 3 depicts a system which will give the desired result described above. System operation is as follows: At "1" the input voltage rises to $+kV_\pi$ and is coupled to "2" through the input capacitor. At "3" the signal has been inverted and amplified to a value that causes the negative-voltage-level sensor to be triggered. At "4" the upper off gate and triggered-current source are enabled. The triggered-current source places a charge at "2" on the input capacitor so that the voltage at that point goes to $-kV_\pi$. The placement of this charge must occur in the shortest possible time since this is the waveform switching interval. This negative voltage is inverted and amplified at "3" to produce a voltage that would trigger the positive-voltage-level sensor except for the earlier action of the upper off gate which disables the positive-voltage-level sensor for a finite length of time sufficient to allow the voltage at "3" to decrease to a value that will not activate the positive-voltage-level sensor. The signal level at "2" is now $-kV_\pi$ and the charge placed on the input capacitor remains there due to the infinite input impedance of the wideband amplifier and the infinite output impedance of the triggered-current sources. The output level of the signal generator at "1" is now $+kV_\pi$ and continues to increase. When it reaches $+3kV_\pi$ the voltage at "2" will have risen to $+kV_\pi$ and the previously described switching action will again be performed. The switching action that occurs when the signal generator goes negative is the same except that the positive-voltage-level sensor is triggered

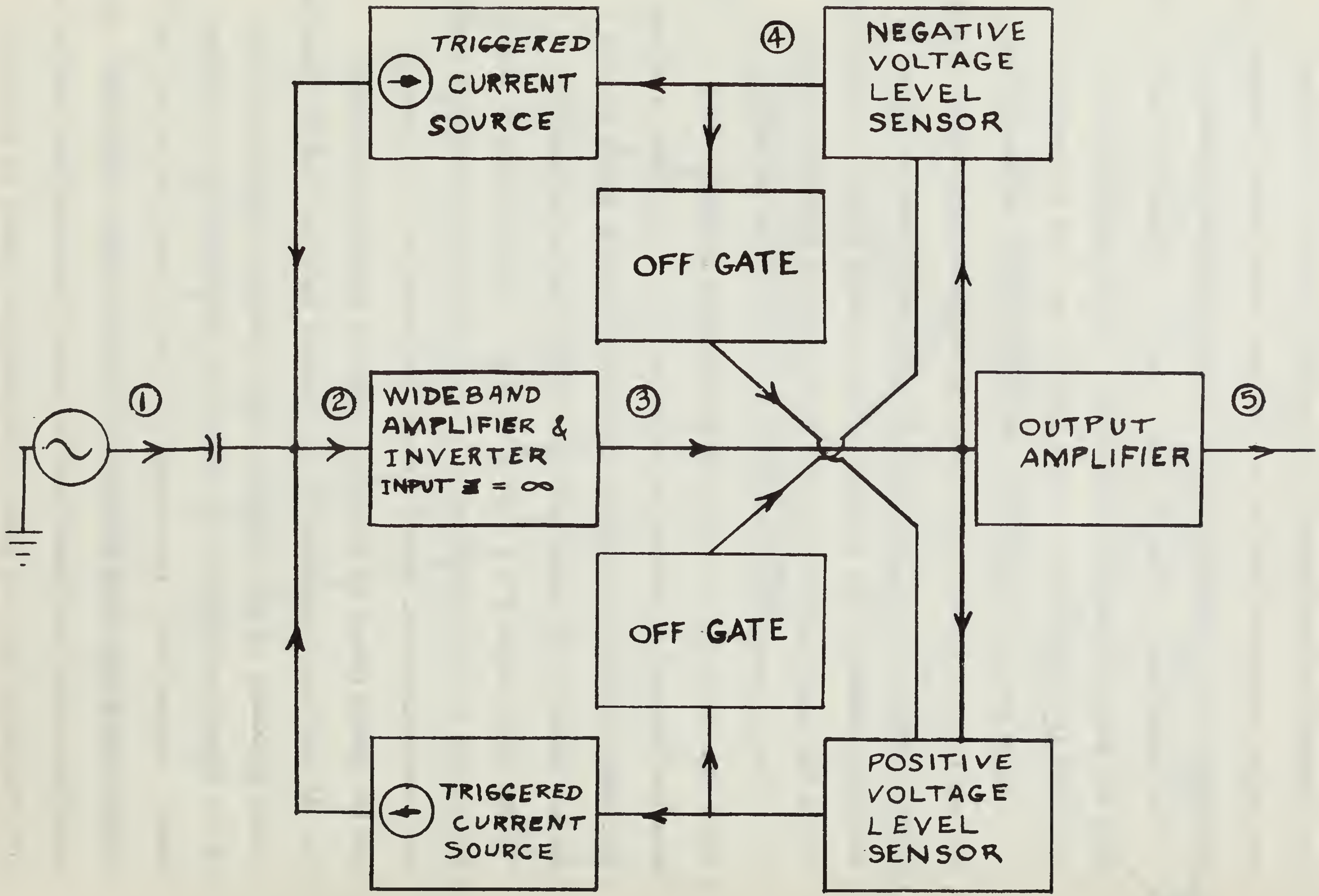



FIG. 3 BLOCK DIAGRAM OF PROPOSED SYSTEM

which results in the triggering of the lower off gate and triggered-current source. The off gates are necessary to prevent self oscillation of the system. Consider the sequence of events if these off gates were not present. Assume the negative-voltage-level sensor has been triggered causing the upper triggered-current source to dump a charge on the input capacitor so that the voltage at "2" is $-kV_\pi$. This voltage is inverted and amplified at "3" causing the positive-voltage-level sensor to trigger which in turn causes the lower triggered-current source to dump a charge from the input capacitor so that the voltage at "2" goes to $+kV_\pi$. This voltage is inverted and amplified at "3" to a value that causes the negative-voltage-level sensor to trigger and the sequence of events is repeated ad infinitum.

Note that if the input signal is increased in amplitude the output will be chopped into more segments. In no case will the output exceed a peak-to-peak swing of $2V_\pi$, and in fact will always vary between $-V_\pi$ and $+V_\pi$ if the peak-to-peak input voltage is equal to or greater than $2kV_\pi$. An implicit assumption is made that the output amplifier amplifies the voltage at "3" to a value that varies between $-V_\pi$ and $+V_\pi$ at "5".

# III. PRACTICAL SYSTEM

Several variations of the basic block diagram shown in Fig. 3 were experimentally attempted before the final circuit shown below in Fig. 4 performed adequately.

The MOSFET amplifier and junction-FET amplifier serve the same purpose as the wideband amplifier with infinite input impedance shown in Fig. 3 with the exception of the dual inversion in the practical system of Fig. 4. Since it is desirable to sense a certain voltage level it was necessary to provide direct coupling up to the voltage-sensing point "3". This necessitated the use of a P-channel and an N-channel FET. The MOSFET was required so that any charge placed on the input capacitor would not be removed through the input impedance of the stage immediately following the input capacitor. The junction-FET amplifier served to isolate the input stage and primarily to restore the output at "3" to the input d-c level. The modified Schmitt triggers replace the voltage-level sensors in Fig. 3. A basic Schmitt trigger [2,3] was modified to provide a means of disabling the Schmitt triggers to prevent self oscillations described in the previous section. The emitter followers were used to prevent loading of the output stage of the modified Schmitt triggers. The 5-μsec monostable multivibrators [4] perform the function of the off gates in Fig. 3. Thus if one Schmitt trigger fires, the other Schmitt trigger is disabled for a period of 5-μsec. The 1-μsec monostable multivibrators, phase splitters, diode-bridge enablers, and the diode bridges with their $kV_{\pi}$-supplies serve the same purpose as the triggered-current sources of Fig. 3. Their operation is as follows: when the

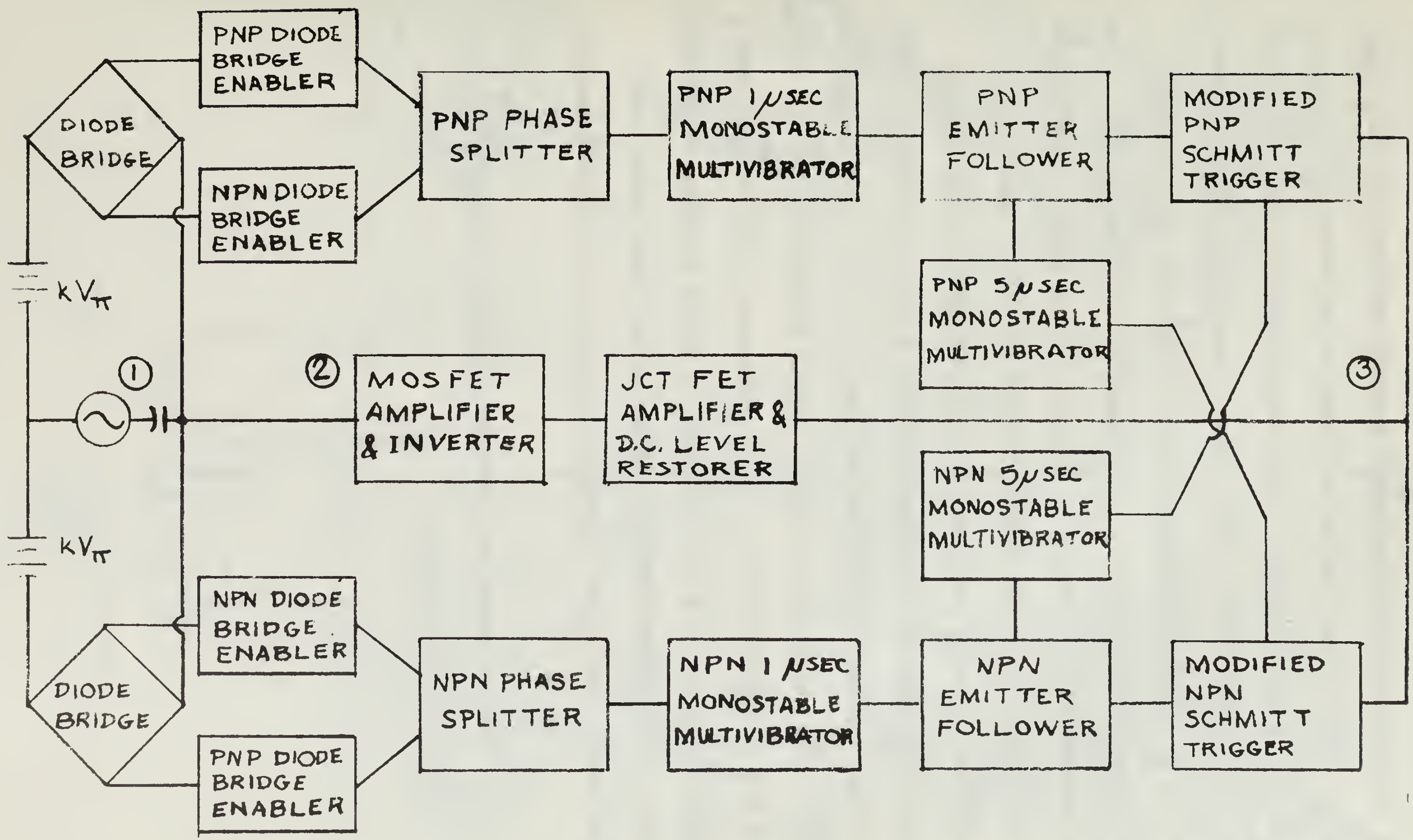



FIG. 4 BLOCK DIAGRAM OF PRACTICAL SYSTEM

1-$\mu$sec monostable multivibrator is triggered, the output is split into a positive and negative pulse of equal amplitude by the phase splitter. These pulses enable the diode bridge through changes in diode-bridge enabler voltages. The diode bridge is normally biased so that all diodes in the bridge have a reverse bias due to operating potentials of the diode-bridge enablers. When the diode bridge is forward biased the supply $kV_{\pi}$ is essentially connected to "2". Thus the voltage at "2" is clamped to $+kV_{\pi}$ or $-kV_{\pi}$ as appropriate through the diode bridge's forward impedance plus the signal generator's output impedance. During the time the diode bridges are not enabled and are reverse biased, the capacitor sees a large impedance through which it must charge, thereby maintaining it's charge for a satisfactory length of time. Note that the input signal generator is in the charging path of the capacitor, hence a signal generator of low output impedance ($50\Omega$) was used. For voice signals it would likewise be necessary to provide a low-output-impedance stage as the final stage in any amplifier chain that might be needed to replace the signal generator shown.

On the following pages is a detailed schematic diagram (Figs. 5 & 6) of the final circuit devised.

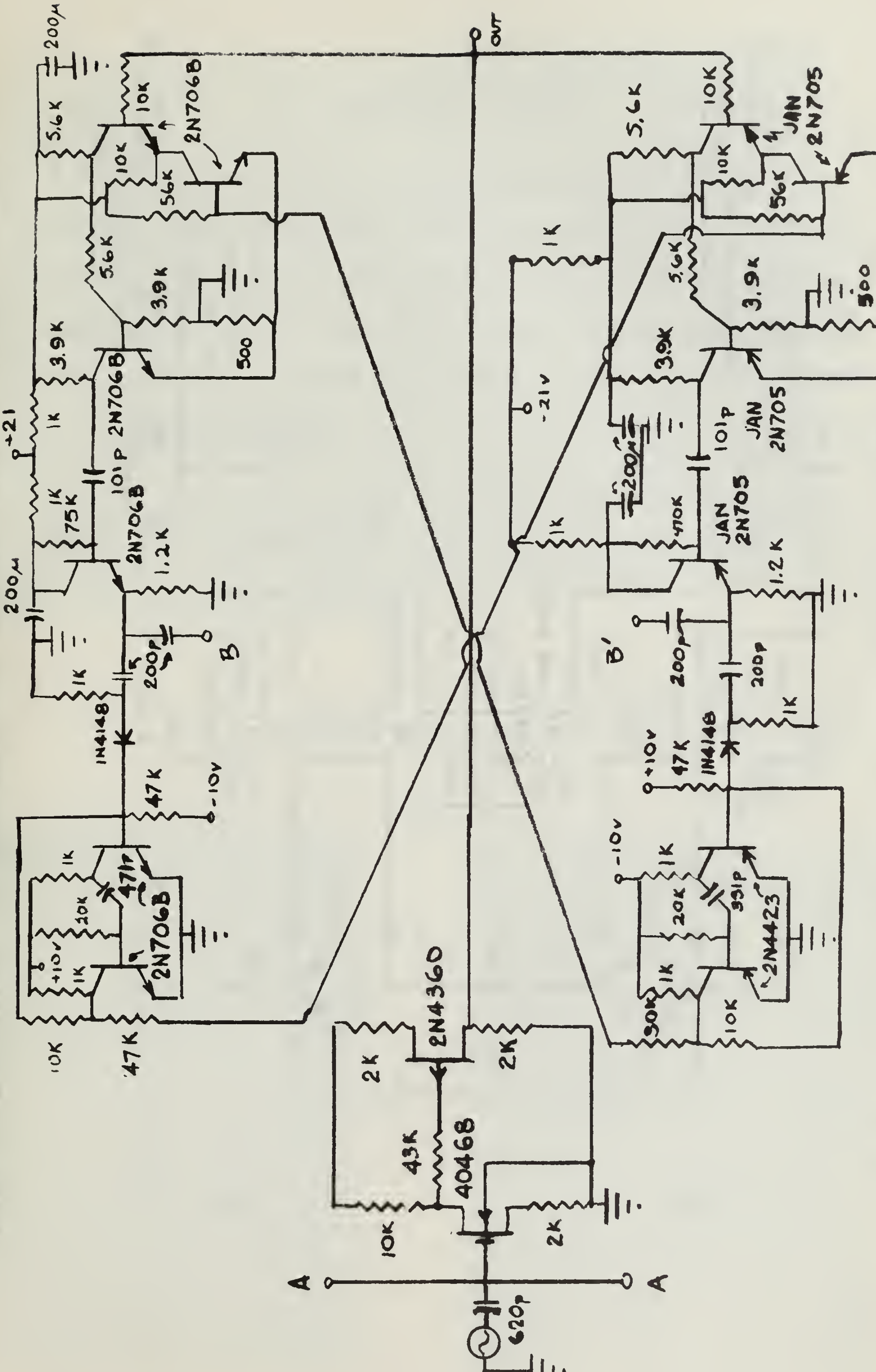



FIG. 5 DETAILED SCHEMATIC OF PRACTICAL CIRCUIT

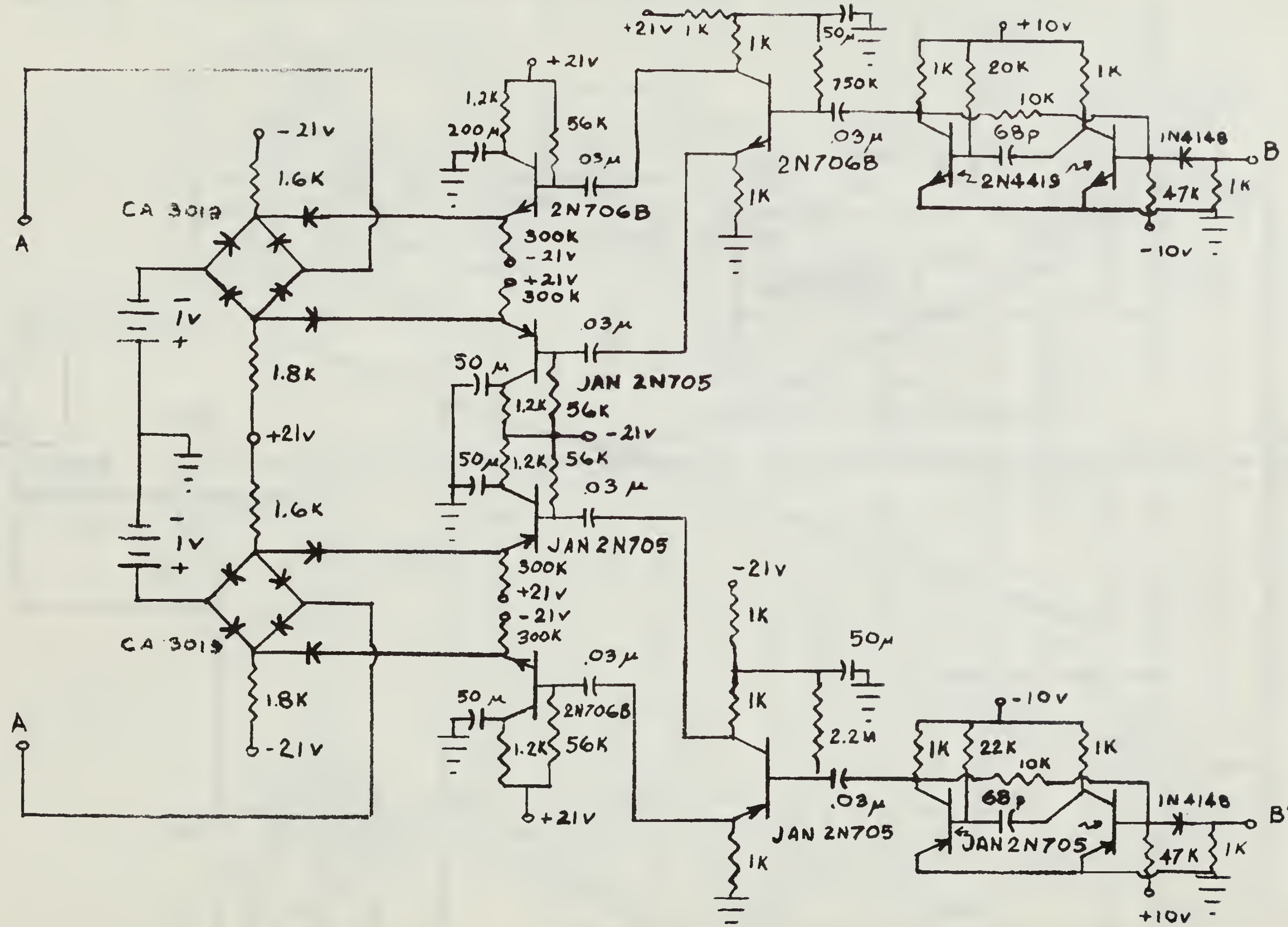



FIG. 6 DETAILED SCHEMATIC OF PRACTICAL CIRCUIT (cont.)

## IV. EXPERIMENTAL RESULTS

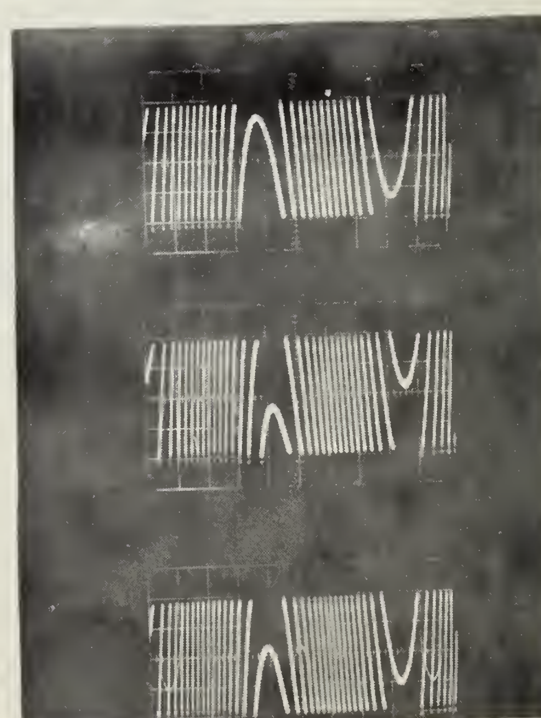
On the following pages are photographs of oscilloscope presentations that resulted from connecting the output of the circuit shown in Figs. 5 & 6 to the input jack of the oscilloscope. The photos are of chopped sine waves of various input frequencies with various amplitude settings of the input signal generator to produce the number of chops shown.

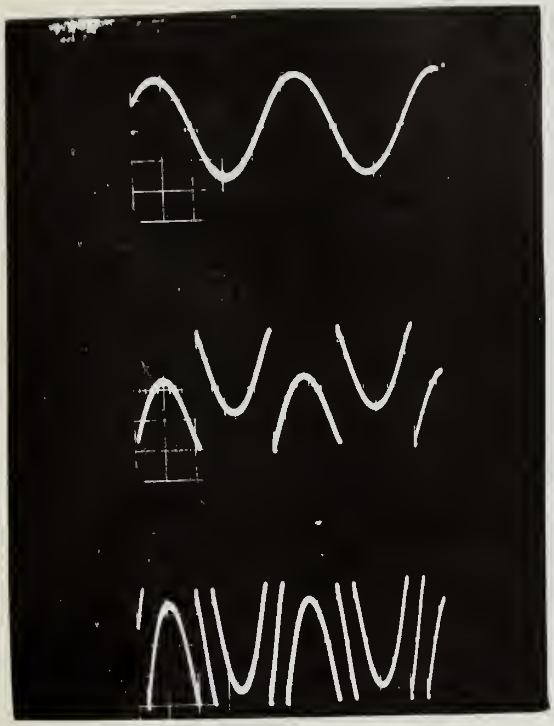
The vertical deflection scale on all photos is 2.0 V/cm. giving a peak-to-peak output of approximately 8.0 volts.

These modified sine waves were taken from the output of the circuit devised (chopper) with an input frequency of 200 Hz. and increasing input amplitudes from top to bottom.

These modified sine waves were taken from the chopper output with an input frequency of 200 Hz. and increasing input amplitudes from top to bottom.

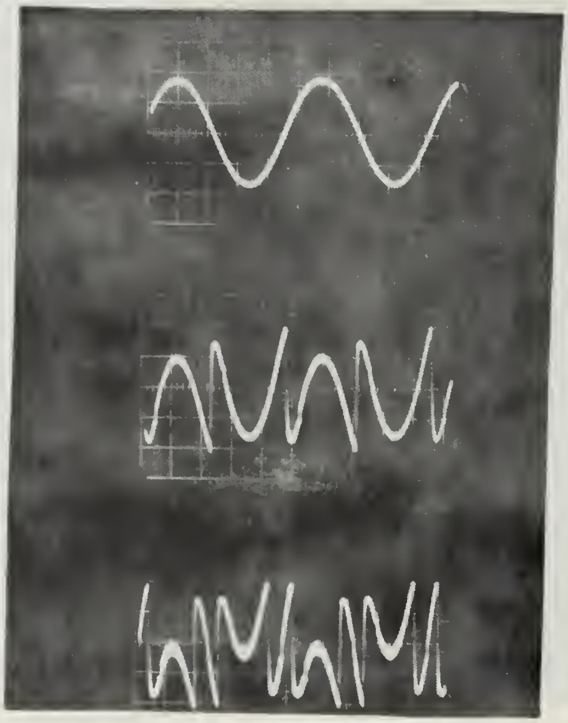
Note: Distortion of the chopped sine wave in the first two frames of the upper photo is due to the signal generator and not due to the chopper.





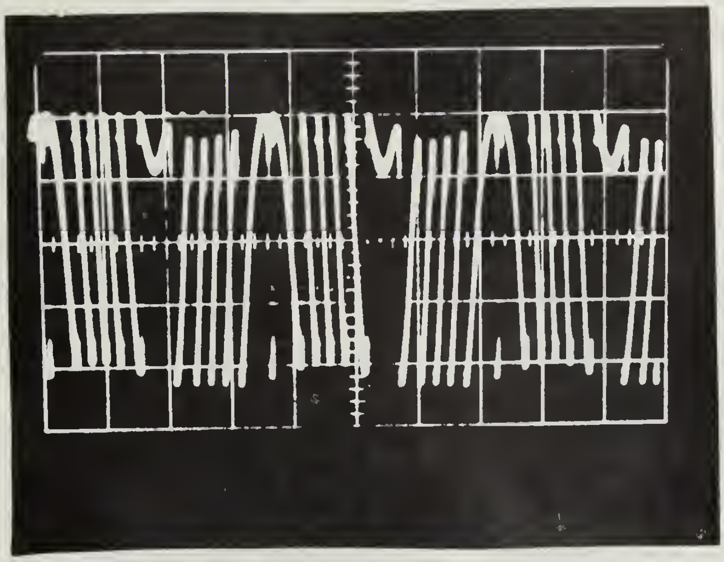
These modified sine waves were taken from the chopper output with an input frequency of 2.0 KHz. and increasing input amplitudes from top to bottom.

These modified sine waves were taken from the chopper output with an input frequency of 2.0 KHz and increasing input amplitudes from top to bottom.



These modified sine waves were taken from the chopper output with an input frequency of 20.0 KHz and increasing input amplitudes from top to bottom.

The original goal was to chop any audio signal as many times as necessary so that the output chopped frequency would be at least 100.0 KHz. For the last frame where 4 chops/cycle have been effected the chopped frequency would be approximately 100.0 KHz although the output waveform is not as close to the desired form as are those with lower chopping frequencies shown earlier.



Input frequency: 5.0 KHz

Chopped frequency: 80.0 KHz approximately

This photo is shown to display the switching action that occurs when the instantaneous slope of the input waveform is zero. A switching point is reached at the positive peak of the input cycle and the normal switching action described earlier, regarding the positive-voltage-level sensor occurs. However, in this case the input signal immediately begins a negative excursion. As soon as the NPN 5-μsec monostable multivibrator returns to its normal off state the negative-voltage-level-sensor immediately fires and a negative spike is generated as shown above.

Since the output signal level as shown in photos on the previous pages was approximately 8 volts peak-to-peak it was necessary to amplify these signals prior to applying them to the helix of a TWT. For the particular TWT used it was necessary to have a peak-to-peak voltage variation of approximately 160 volts to produce a $2\pi$ phase shift in the output signal. This amplification was achieved through the use of an amplifier used by Fletcher in a related thesis [5]. On the following page is a diagram, Fig. 7, of the microwave setup similar to that used by Fletcher to observe the output spectrum of the TWT. This microwave circuit was necessitated by the unavailability of an extremely stable X-Band signal source for use as the radio-frequency carrier. Also the X-Band spectrum analyzer that was available did not have frequency dispersion characteristics sufficient to show the individual spectral lines. The lower-frequency spectrum analyzer used had this capability; however, it too had a deficiency in that it could not simultaneously display the complete spectra of signals with spectral widths greater than 30 KHz. Note that in the circuit shown stability of the X-Band source was of no consequence since the translated carrier frequency transmitted to the spectrum analyzer exactly equals the frequency of the 30-MHz source. Any frequency drift or incidental frequency modulation in the X-Band source is cancelled in the output. Hence the stability requirement was met through the use of a relatively stable, lower-radio-frequency source. On the pages following Fig. 7 are photos of the output spectrum after translation to a frequency of 30 MHz. Superimposed on three of the photos are the chopped sine waves before amplification that produced the modulation.

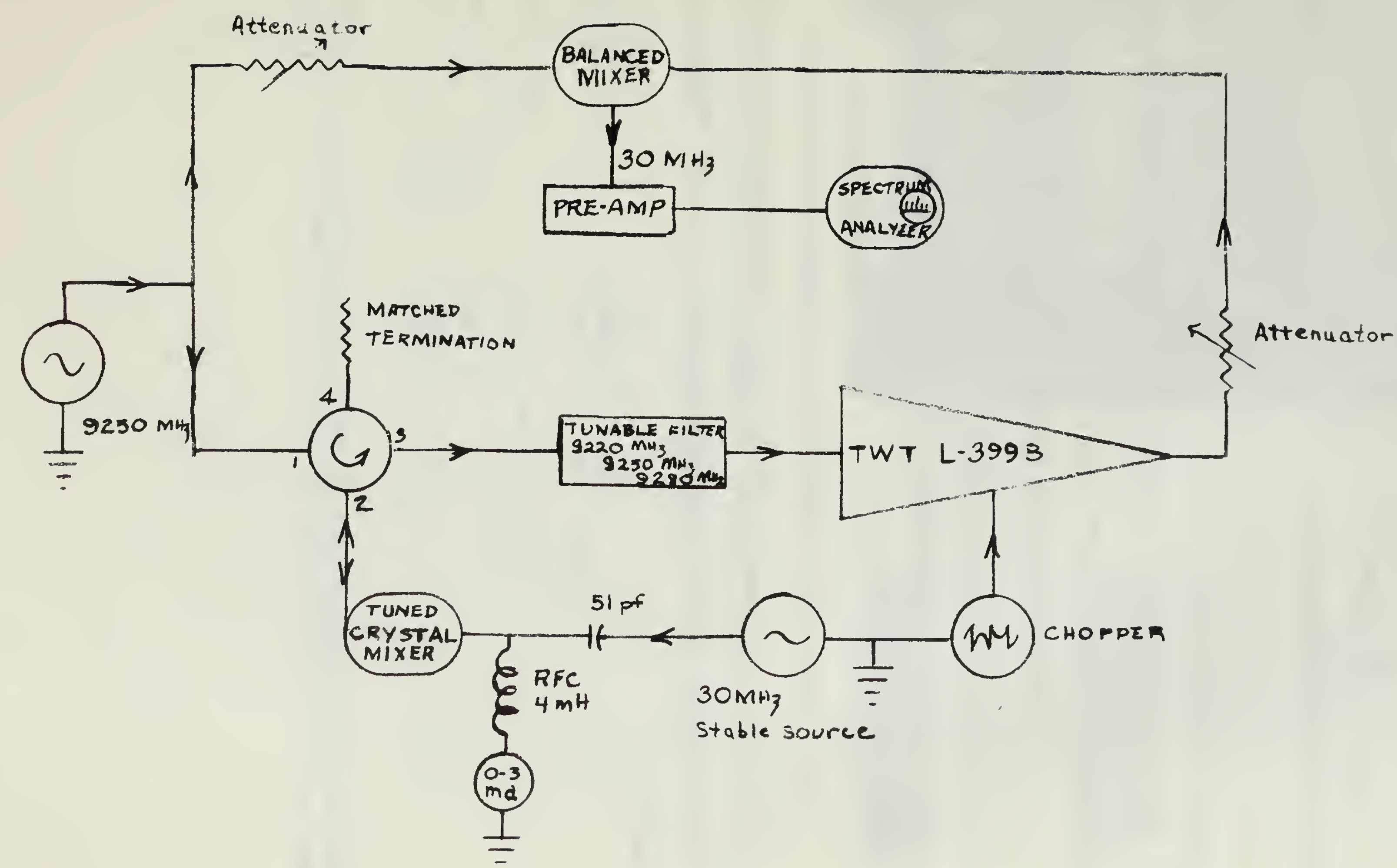
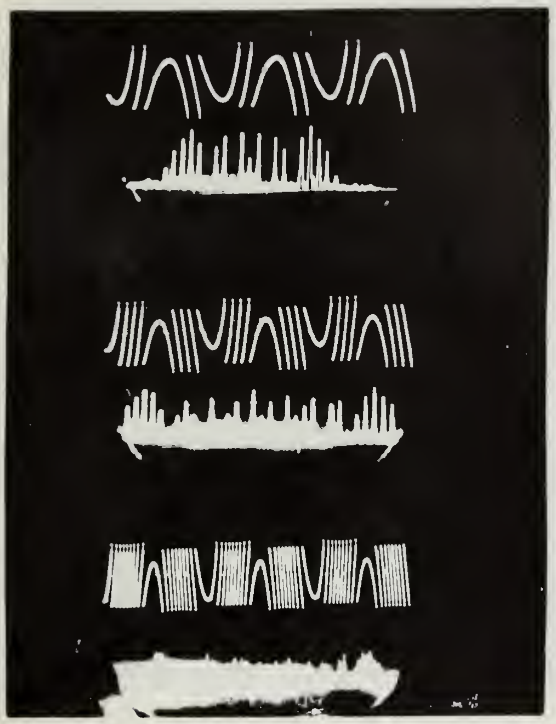
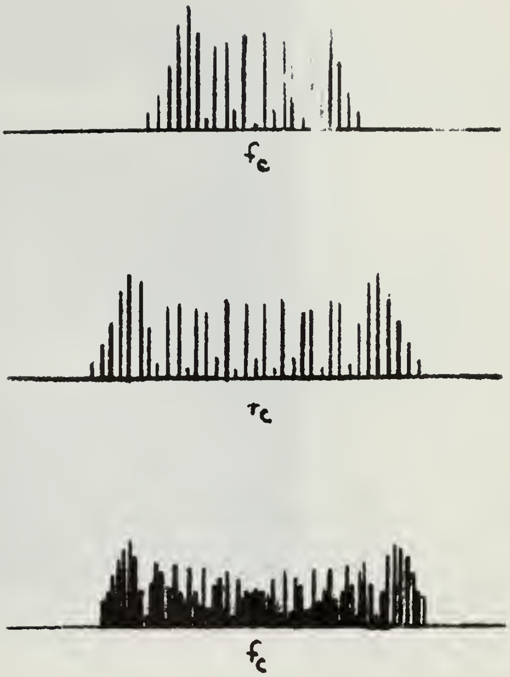



FIG. 7 MICROWAVE CIRCUIT USED TO VIEW OUTPUT SPECTRUM OF TWT

A quick glance at these photos shows that the more times the input signal is chopped, i.e., greater input amplitude, the greater the frequency deviation of the output carrier, as predicted. Additionally, to show that the output spectra obtained with the system developed here contain essentially the same sidebands as would be obtained from a sine wave used to phase modulate a carrier by the specified high modulation index, computed spectra are shown immediately adjacent to the measured spectra. While there is not exact correlation between the measured spectra and perfect theoretical spectra, the correlation is sufficient to justify the earlier statements regarding achievement of broadband phase modulation via chopping of the input signal.

If the input to the chopper were a complex waveform such as voice signals the intelligence could be readily recovered by applying the phase-modulated radio-frequency signal to a receiver employing a phase detector.






Modulating frequency into chopper -- 500 Hz.

| Chops/cycle | Modulation Index |
|---|---|
| Frame A --- 4 | Frame A --- 8.65 |
| Frame B --- 8 | Frame B --- 14.7 |
| Frame C --- 18 | Frame C --- 29.8 approx. |

| Spectrum Width | Significant Sidebands |
|---|---|
| Frame A --- 9.0 KHz | Frame A --- ± 9 |
| Frame B ---15.0 KHz | Frame B --- ± 15 |
| Frame C ---30.0 KHz | Frame C --- ± 30 |

Comments: All spectra except the one on the following page resulted from tuning the filter preceding the TWT shown in Fig. 7 to the upper sideband.

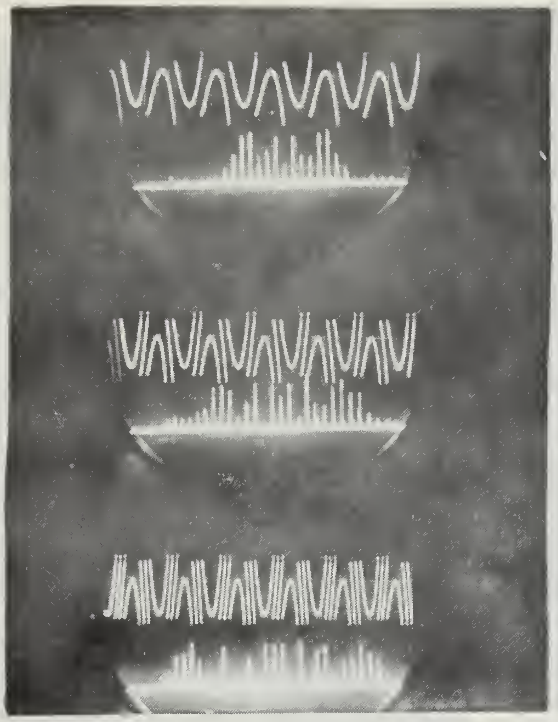
Note: For theoretical spec
see preceeding page.

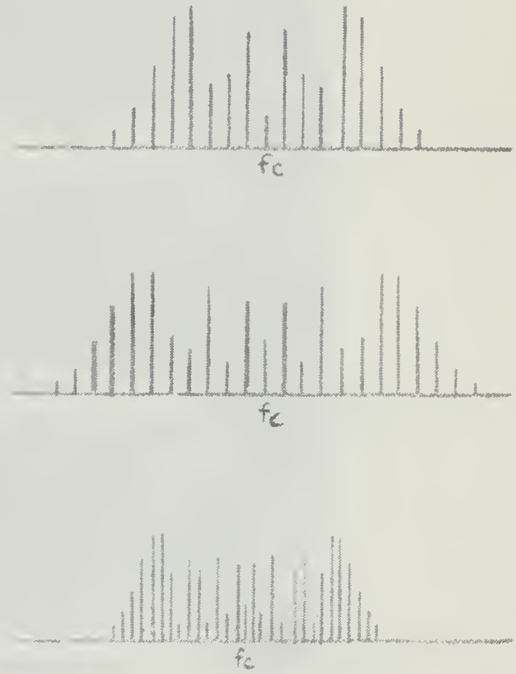
Modulating frequency into chopper -- 500 Hz.

| Chops/Cycle | Modulation Index |
|---|---|
| Frame A --- 18 | Frame A --- 29.8 |
| Frame B --- 4 | Frame B --- 8.65 |
| Frame C --- 8 | Frame C --- 14.7 |

| Spectrum Width | Significant Sidebands |
|---|---|
| Frame A --- 30.0 KHz | Frame A --- $\pm$ 30 |
| Frame B --- 9.0 KHz | Frame B --- $\pm$ 9 |
| Frame C --- 15.0 KHz | Frame C --- $\pm$ 15 |

Comments: The above spectra resulted from tuning the filter preceding the TWT to the lower sideband.

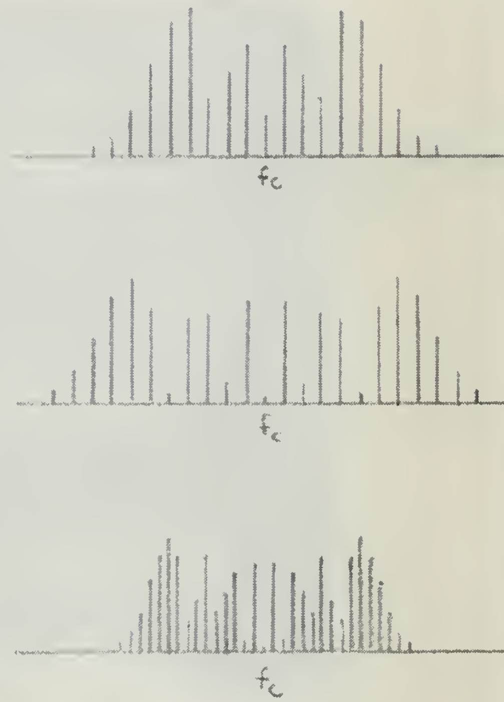





Modulating frequency into chopper -- 1.0 KHz.

| Chops/Cycle | Modulation Index |
|---|---|
| Frame A --- 2 | Frame A --- 5.8 |
| Frame B --- 4 | Frame B --- 8.1 |
| Frame C --- 6 | Frame C ---11.3 |

| Spectrum Width | Significant Sidebands |
|---|---|
| Frame A --- 12.0 KHz | Frame A --- ± 6 |
| Frame B --- 16.0 KHz | Frame B --- ± 8 |
| Frame C --- 22.0 KHz | Frame C --- ±11 |




Modulating frequency [illegible] chopper -- 1.0 KHz.

Chops/Cycle

Frame A --- 2
Frame B --- 4
Frame C --- [illegible]

Modulation Index

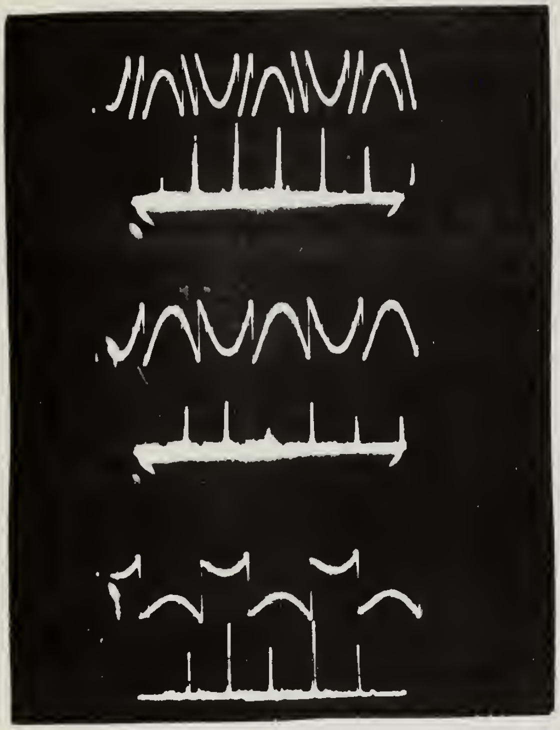
Frame A --- [illegible].9
Frame B --- 8.6
Frame C --- 11.85

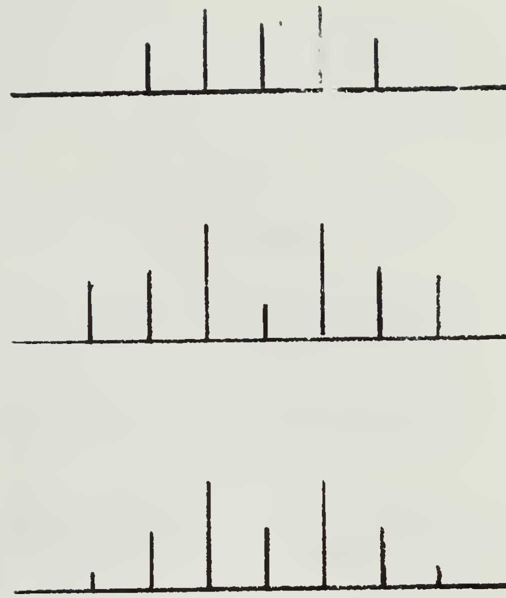
Spectrum Width

Frame A --- 12.[illegible] KHz
Frame B --- 1[illegible].0 KHz
Frame C --- 24.0 KHz

Significant Sidebands

Frame A --- ± 6
Frame B --- ± 9
Frame C --- ± 12





Modulating frequency into chopper 5.0 KHz.

Chops/Cycle

Frame A --- 4
Frame B --- 2
Frame C --- 2

Modulation Index

Frame A --- 7.9
Frame B --- 5.8
Frame C --- 1.9

Spectrum Width

Frame A --- not all shown
Frame B --- not all shown
Frame C --- 20.0 KHz.

Significant Sidebands

Frame A --- not shown
Frame B --- not shown
Frame C --- $\pm$ 2

Comments: These spectra show the deficiency of the spectrum analyzer used as mentioned earlier when discussing spectrum width.

Frame C displays an interesting facet of the chopper. Although the input signal level would not normally be sufficient to cause the chopping shown it was possible by first increasing the input signal level to a value sufficient to cause 2 chops/cycle (which is the minimum number of chops possible). Then by gradually decreasing the input amplitude below the normal chopping level the unusual waveform in Frame C was obtained. This waveform is a direct consequence of the direct coupling that exists between the input MOSFET amplifier, the junction FET amplifier and the modified Schmitt triggers.

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13. ABSTRACT

If a sawtooth wave is used as the modulating signal for the accelerating voltage of a traveling-wave tube, the input radio-frequency signal is phase modulated to produce a frequency shift in the output signal. The use of a sine wave as the modulating signal results in the familiar sine-wave-phase-modulated carrier as the output; however, it is difficult to obtain a large phase-modulation index without objectionable amplitude modulation. Specifically, if the peak-to-peak voltage swing of the modulating wave produces a phase shift of more than $2\pi$ radians in the radio-frequency carrier, the amount of amplitude modulation becomes severe. To overcome this severe amplitude modulation and yet produce large phase deviations in the carrier signal, i.e., broadband phase modulation, it is possible to chop the modulating waveform just as a ramp is chopped into a sawtooth. This thesis presents a design for so modifying the modulating signal be it sine wave, sawtooth, exponential or complex, and shows the resultant output spectrum of a traveling-wave tube whose carrier has been modulated by a chopped sine wave.

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| Limited-Continuous Phase Deviation | | | | | | |
| Broadband High Input Impedance Amplifier | | | | | | |
| Modified Serrodyne Modulation | | | | | | |
| Chopped Waveforms | | | | | | |